AN INSTANTANEOUS AC VOLTAGE AND WAVEFORM STABILIZER

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AN INSTANTANEOUS AC VOLTAGE AND WAVEFORM STABILIZER

by

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PREFACE

Today there are many uses for large quantities of AC power of constant voltage amplitude and pure voltage waveform. This paper contains the details of an investigation into an inexpensive and reliable method for providing a large amount of stabilized AC power. In the systems described, the major portion of the power supplied to the load comes from the power lines, with the stabilizer supplying or absorbing power as the input line voltage may require. Although all of the investigation was made for a power line frequency of 60 cycles per second, there is very little reason why similar devices could not be employed for other power line frequencies, in particular, 400 cycles per second. In the discussion that follows, mention is made of all components that would require changing for a power line frequency different from 60 cycles per second.

The major portion of the preliminary work on the stabilizer was accomplished during the writer's commercial laboratory term spent at the Hewlett Packard Company, Palo Alto, California. The author wishes to acknowledge the helpful assistance provided by Dr. Barney Oliver, Mr. Brunton Bauer, who suggested the subject, and Mr. Bruce Whooley of the Hewlett Packard Company. Grateful acknowledgment is also made of the editorial criticism and comments by Professor J. J. Downing, Jr., of the Naval Postgraduate School.

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SUMMARY

A brief general discussion is given of the uses for which precision AC voltage and waveform stabilizers are required. This discussion is followed by a description of the four types of stabilizers that have been designed and the good and bad features of each. Choosing the buck-boost control system as the preferred type, the basic theory of such a system in its various forms is discussed. The theory is given for obtaining a constant voltage reference source by heavy clipping of the input sine wave voltage. Circuit diagrams, analysis, and actual performance of the stabilizer in its three forms are given. Also included are recommended systems for improving the reference voltage stability and for obtaining a great increase in the power that the stabilizer can control.

I. INTRODUCTION

Although many electrical devices consuming AC power will function reasonably well with input voltage variations of plus or minus 15%, there are some purposes for which an AC voltage of constant amplitude and pure sinusoidal waveform is required. Typical applications are constant light sources in the field of applied optics, the calibration of AC instruments, X-ray diffraction cameras, high-precision selsyns, and microwave bridges.

Depending upon the generating source and the power line loads, total harmonic voltage distortions of 5 to 15% are not uncommon--particularly in locations where heavy electrical machinery is operated. Much effort has been expended by the power companies in improving the voltage stability of the AC power delivered to their customers--not only because of the complaints of the more observant customers, but also because of the additional revenue derived from maintaining a high average line voltage at the consumer delivery point. Unfortunately, very little can be done to improve the output waveform of an installed generator except to attempt to keep the loads on the three phases of the alternator balanced at all times. Considerable unbalance of the loads on each of the three phases may result in the circulation of third, fifth, and seventh harmonic currents, with a resultant high total harmonic distortion content for the AC power delivered to the ultimate consumer.

Since the average consumer can do very little about the voltage stability and harmonic content of the AC power delivered to him on the power lines, his only recourse is to try to devise some means of reducing the voltage amplitude variations and total harmonic content of that AC power.

This reasoning has led to the development of several different AC voltage and waveform stabilizers that serve only to modify the delivered AC power. Although some of the development has been done in the United States, the major portion of the earlier work was done in England.

The various types of stabilizers discussed by Maddock[1] are:

(1) Saturable transformers

(2) Series-reactance control

(3) Buck-boost control

(4) Transductor

The saturable transformer stabilizer suffers from waveform distortion, frequency sensitivity, and provides voltage amplitude stability of only about 1%, although it offers a means of controlling large amounts of power. The transductor stabilizer (a constant current device) suffers from severe waveform distortion, but provides regulation of about 0.1% for up to several KVA of controlled power. The series-reactance stabilizer provides a fairly good output voltage waveform with an amplitude stability of 0.15%, and has been used to control up to 5 KVA, but it requires a special sensing diode that can be obtained only from the factory of the company manufacturing the device (Sorensen and Superior). On the other hand, buck-boost stabilizers have been built that provide a very good output voltage waveform ($<$ 1% distortion), a voltage amplitude stability of 0.01%, and a controlled power of 1 KVA. In addition, the buck-boost stabilizer offers the advantage of almost instantaneous correction for changes in input voltage and/or waveform, whereas the other systems mentioned require from 10 to 20 cycles of the line frequency to adjust to shifts in input voltage. With proper design it is also possible to have the buck-boost stabilizer work over a wider

range of input frequency than is possible with any of the other types mentioned above, most of which are quite frequency sensitive because of the very principle of their operation.

Since the buck-boost system offers so many advantages over the other types of stabilizers, further study of this method of voltage and waveform stabilization seems to be indicated. The basic buck-boost block diagram is indicated in Figure 1.

Defining the input and output voltages as E_i and E_o respectively, and δE_o as the change in E_o resulting from a change δE_i in input voltage E_i, the fractional changes are $\delta E_i/E_i$ and $\delta E_o/E_o$, and the Stabilization Ratio is:

$$S \equiv \frac{\delta E_i/E_i}{\delta E_o/E_o} = \frac{\delta E_i}{\delta E_o} \frac{E_o}{E_i}$$

The value of S, always greater than unity, indicates the improvement in stability of the output voltage over the input voltage.

A second characteristic that is of great importance in a stabilizer is its effective internal resistance, where

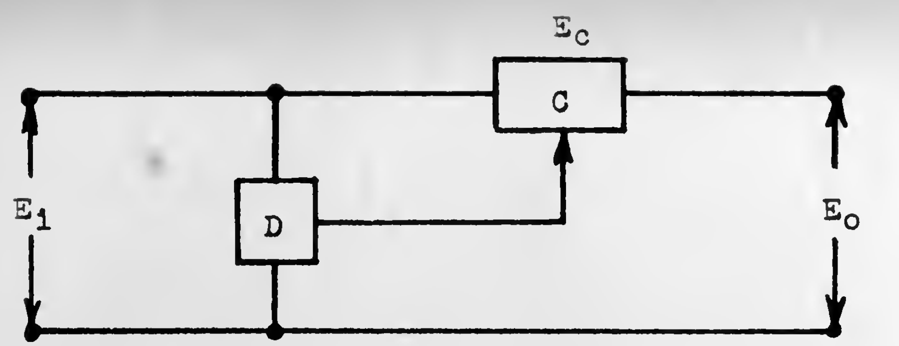
$$R_o \equiv \delta E_o/\delta I_o$$

The functional performance of a stabilizer is completely defined by the values of S and R_o.

There are at least three possible configurations that the buck-boost system may take:

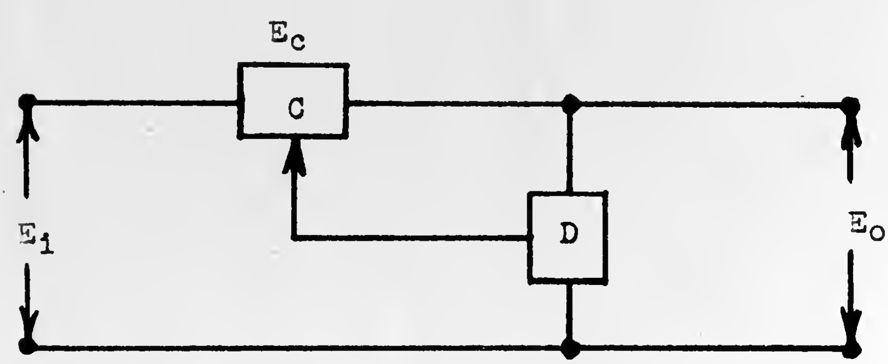
(1) Detector across the output

(2) Detector across the input



$$S \equiv \frac{(\delta E_i)(E_o)}{(\delta E_o)(E_i)} = \frac{1}{1 - x_i A_i}$$

(a) Detector across stabilizer input



$$S \equiv \frac{(\delta E_i)(E_o)}{(\delta E_o)(E_i)} = 1 + x_o A_o$$

$$R_o = \frac{-R'}{S} = \frac{-R'}{1 + x_o A_o}$$

(b) Detector across stabilizer output

Figure 1. Basic Block Diagram For A Buck-Boost Stabilizer

(3) Detectors across the input and output

Patchett[2] discusses the various advantages and disadvantages of having the detector in positions (1), (2), and (3).

For correction in output voltage due to changing load and/or changing input voltage, the best position for the detector is across the output of the stabilizer. For this position, Maddock[1] derives the relations:

$$S = 1 + x_o A_o$$

$$\text{and } R_o = \frac{-R'}{1 + x_o A_o} = \frac{-R'}{S} \; *$$

where: X_o = the fractional part of the output voltage taken by the detector and balanced against a reference voltage E_r

A_o = total amplification given the error voltage resulting from $(x_o E_o - E_r)$

R' = actual internal resistance of the stabilizer

All of the buck-boost stabilizers employ essentially the same general principle of operation. The differences between the various successful stabilizers are essentially in the methods employed to obtain an error voltage that may be amplified and placed in series with the input voltage to keep the desired constant output voltage of pure waveform Obviously this procedure involves the possibility of hunting in the output voltage, since a change in output voltage is required before a correction voltage

* Appendix I

can be developed. The methods by which the various stabilizers obtain an error voltage and prevent the output voltage from hunting are described in the following chapters.

II. NON-LINEAR ELEMENT AC BRIDGES

A non-linear element AC bridge may be used to furnish a constant output reference voltage or to furnish an output error voltage whose magnitude and phase depend upon the change in the bridge input voltage. The theory involved in the latter usage, to provide an error voltage, is discussed in the following paragraphs.

Non-linear elements in general can be divided into two classes. Class I includes those elements such as tungsten-filament lamps and barretters whose resistance increases with an increase in current. Class II includes those elements such as silicon carbide (Thyrite) and carbon-filament lamps whose resistance decreases with an increase in current. Although the Thermistor exhibits both positive and negative slopes on its voltage-current characteristic, it is generally considered to fall in Class II because it exhibits a negative slope over the major portion of its characteristic.

To avoid the changes in average or peak value caused by the presence of harmonic components in the output voltage, Patchett(2) shows that it is highly desirable to maintain the RMS value or the fundamental component of the output voltage constant in high precision stabilizers. Figure 2 shows a bridge with resistor arms R_1, R_2, R_3, and R_4. One or more of these resistor arms must employ a Class I or Class II non-linear element in order for the bridge to be voltage-sensitive. Only those non-linear elements whose action is not instantaneous can be employed satisfactorily in this type of bridge, since the resistance variation should not follow the voltage variation throughout a cycle.

In all of the satisfactory non-linear devices: carbon and tungsten filament lamps, barretters, saturated diodes, and Thermistors, the non-linear action is brought about by temperature changes and there is, consequently,

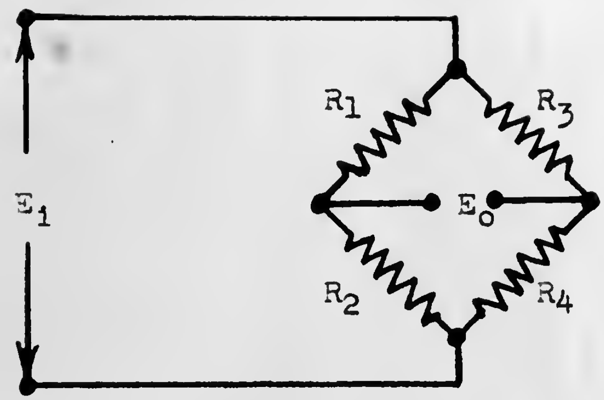


Figure 2. Basic Resistor Bridge

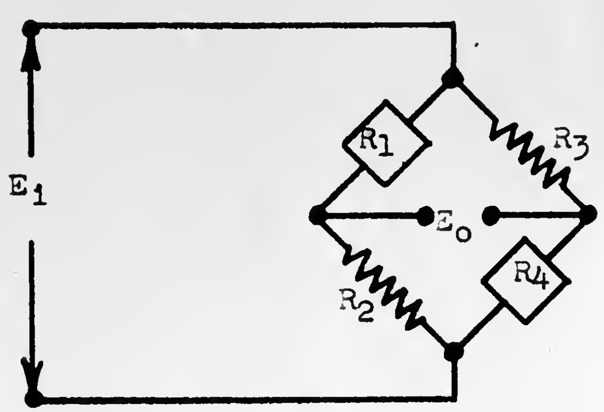


Figure 3. Voltage-Sensitive Bridge Employing
Two Non-Linear Elements

a time lag between a voltage change and the resulting resistance (or emission) change. When the voltage-sensitive bridge is used on AC of relatively low frequency, the small cyclical variation in temperature results in cyclical changes in the resistance of the non-linear element. Normal balancing of the bridge leaves an unwanted voltage output from the bridge which Patchett[2] and Cunningham[3] have shown to consist of two equal amplitude voltages:

(1) One voltage of three times the line frequency

(2) One voltage of the line frequency but in quadrature with the line voltage

Over a limited range of voltage, the characteristics of non-linear elements follow the relation: $E = kI^n$, where k and n are constants determined by the particular element considered. For the arrangement of the non-linear elements and linear resistors as shown in Figure 3 and operated at the E_i for which $E_o = 0$, Patchett[2] gives the Figure of Merit $\left(\frac{dE_o}{dE_i}\right)$ as:

$$F = \frac{n-1}{n+1} \, *$$

Typical values of n are: 2.0 for gas-filled tungsten-filament lamps, 0.78 for carbon-filament lamps, 5 to 40 for barretters, and -0.5 for Thermistors. Thus for two gas-filled tungsten-filament lamps, the theoretical Figure of Merit is:

$$F = \frac{2 - 1}{2 + 1} = 0.33$$

Although it is desirable to have as large a Figure of Merit as possible (n either greater or smaller than one), even more important is the ratio of wanted to unwanted voltage output from the bridge. On the basis of his analysis, Patchett[2] reached the following conclusions:

(1) The amplitude of the unwanted voltages increases with bridge voltage

(2) Increasing the lamp wattage decreases the unwanted voltages

* Appendix II

(3) The unwanted voltage output is larger from gas-filled and single-coil lamps than from vacuum and double-coil lamps

(4) Carbon-filament lamps give a smaller unwanted voltage output than tungsten-filament lamps

(5) The unwanted voltages decrease with an increase in frequency

If the gas-filled tungsten-filament lamp is operated below its normal voltage, it offers the advantages of long life and relative immunity to changes in ambient temperature. When used in a voltage-sensitive bridge, however, it gives a relatively low Figure of Merit, a large unwanted voltage, and vibration instability.

A barretter bridge offers a larger Figure of Merit and a larger ratio of wanted to unwanted voltages than a lamp bridge. However, the barretter bridge suffers greatly from changes in ambient temperature and air circulation, and also requires an appreciable time to reach its new equilibrium condition after a change of voltage.

For a voltage-sensitive bridge consisting of three linear arms and one non-linear arm, Patchett[4] derives the relation:

$$\text{Figure of Merit} = \frac{p(n - 1)}{(np + 1)\,(p + 1)}$$

where $p = R_2/R_1$

For $p = 1$, this gives a figure of merit of one-half that of the bridge containing two non-linear elements. It is obvious from this expression that if $np = -1$, the theoretical Figure of Merit of the bridge becomes infinite. It is impossible to make $np = -1$ by using such non-linear elements as tungsten, for which $n = +2$, but for a Thermistor with $n = -0.5$ we get $np = -1$ when $p = R_2/R_1 = 2$. Thus for the Thermistor bridge, because of its extremely high Figure of Merit, we can obtain a much higher ratio of wanted to unwanted output voltage than from the lamp bridge. Since the operating temperature of the Thermistor is normally only slightly above the ambient

temperature, the Thermistor bridge operation is affected appreciably by changes in ambient temperature. Patchett[2] overcame this latter difficulty by a temperature compensation circuit and proceeded with the design of his stabilizer, utilizing a Thermistor bridge.

In reviewing all of the articles that could be found in the literature on the subject of AC stabilizers, it appeared that at least three phases of the problem were still open to investigation:

(1) The use of a two lamp bridge to provide a constant voltage reference source of adjustable phase relative to the output voltage

(2) The use of a two lamp bridge to provide a null-seeking detector

(3) Heavy clipping of a sine wave voltage, of adjustable phase relative to the output voltage, with subsequent extraction and amplification of the fundamental component of the resultant square wave to provide a stable reference voltage of pure waveform

The two types of lamp bridges are discussed in the balance of this chapter, and the clipping system is covered in the next chapter.

1. Constant output lamp bridge

Since very little information is available in the literature on the characteristics of American lamps of various wattage ratings, curves were run for three different types of General Electric lamps: 3 watt-120 volt, 6 watt-120 volt, and 10 watt-250 volt. All of these lamps are used in various items of Hewlett Packard test equipment, and it seemed that one of the lamps might be a suitable inexpensive and trouble-free non-linear element to use in a constant output lamp bridge. Figure 4 is a plot on log-log paper of lamp resistance versus bridge voltage. It will be noted that the actual lamp voltage is only half of the applied bridge voltage, since $R_1 = R_2$. Although the curves were run for various amplitude 60 cycle input voltages, no appreciable change in lamp resistance was noted at slightly higher or lower frequency voltages.

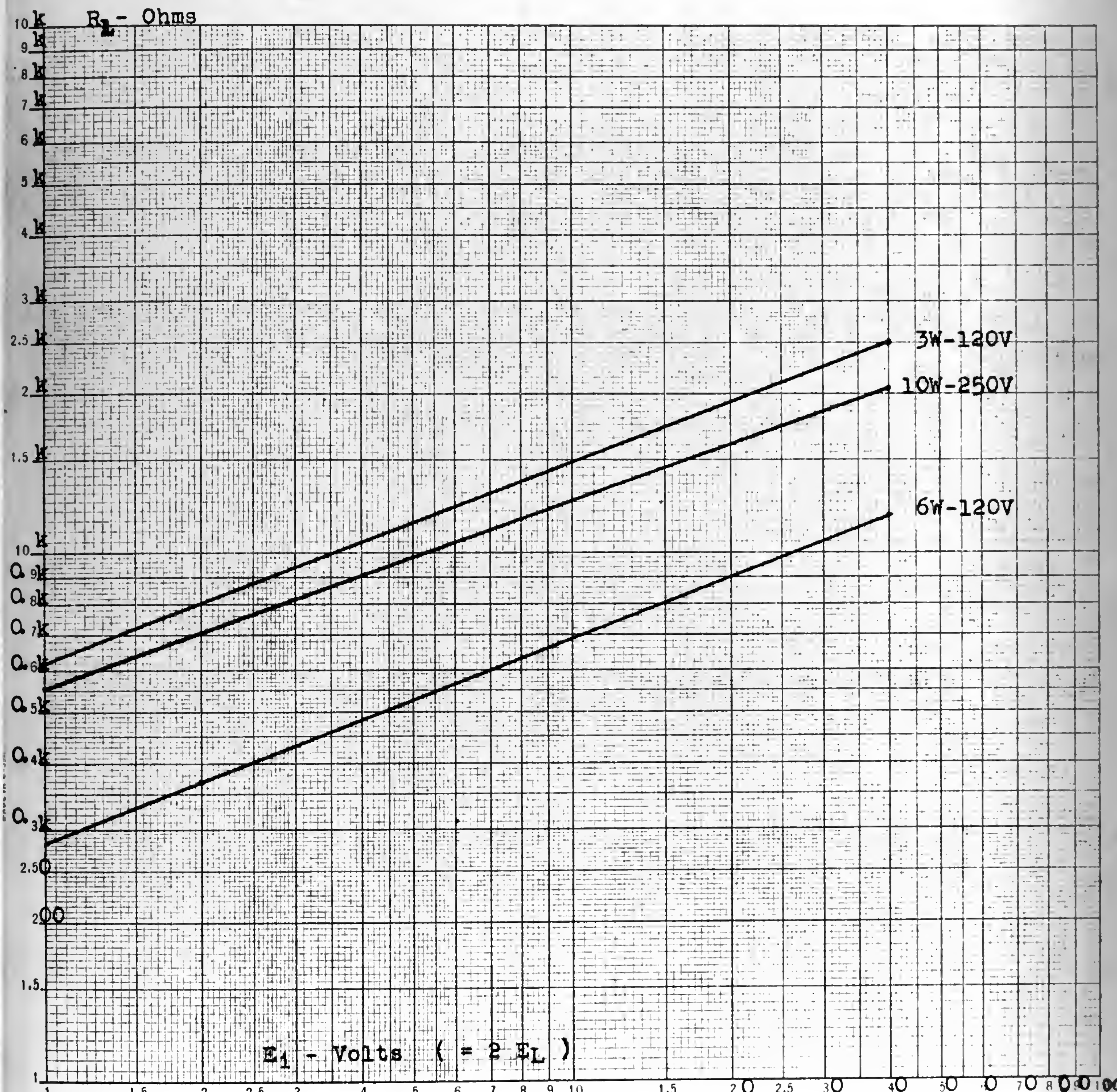


Figure 4. Characteristics Of Three General Electric Lamps

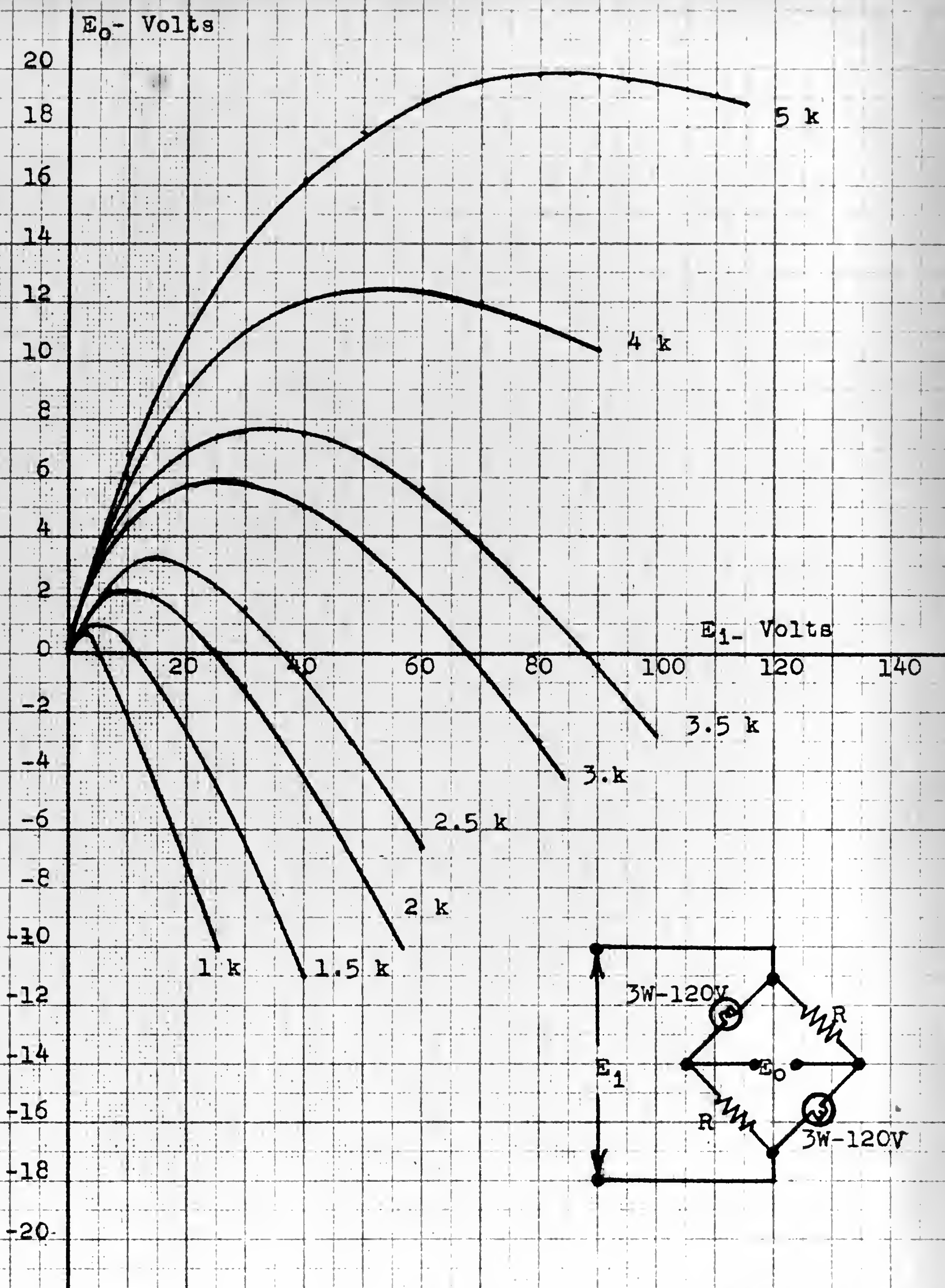


Figure 5. E_o Versus E_i For A 3W-120V Lamp Bridge

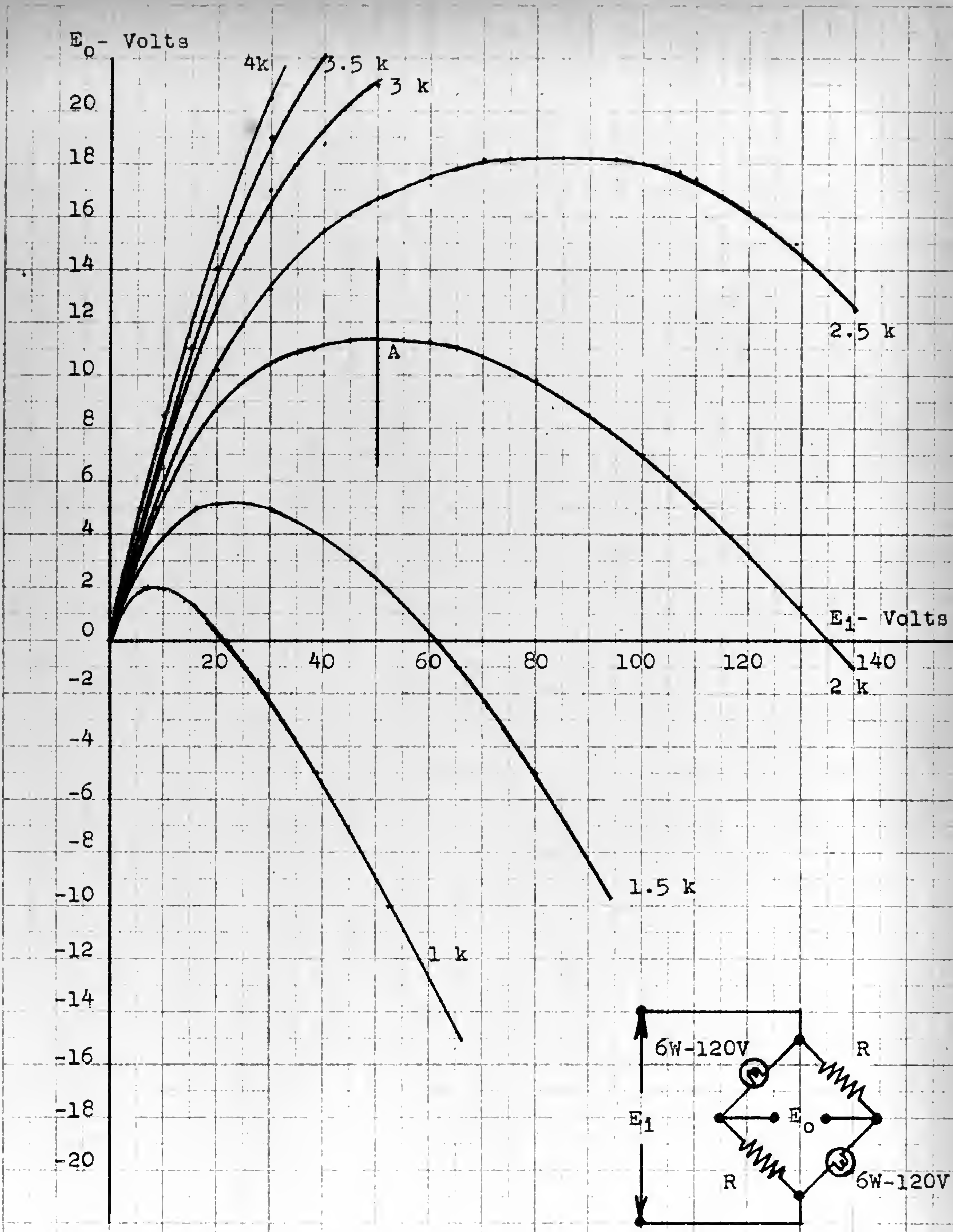


Figure 6. E_o Versus E_1 For A 6W-120V Lamp Bridge

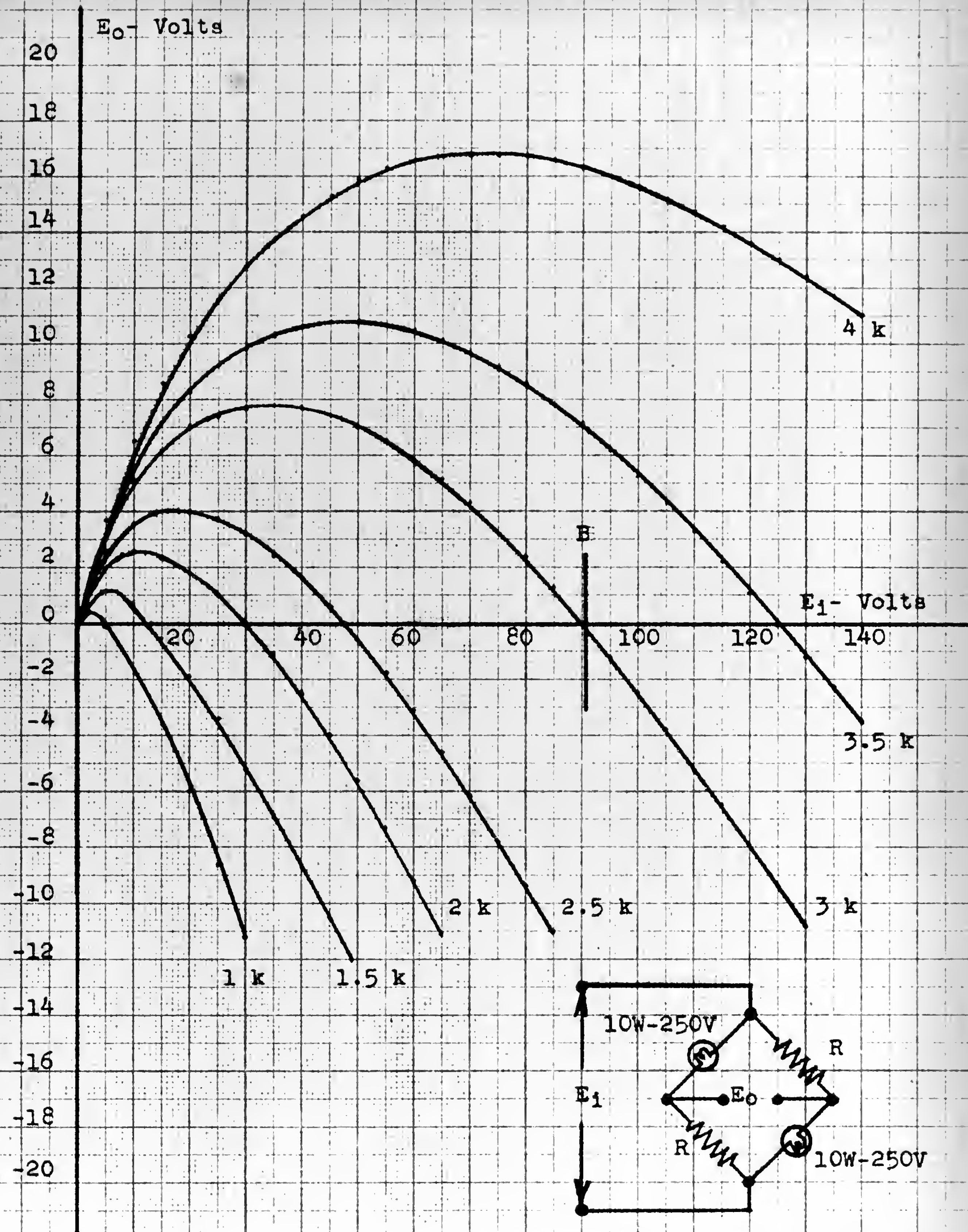


Figure 7. E_o Versus E_i For A 10W-250V Lamp Bridge

Each set of lamps was then connected in a bridge circuit, and curves run of output voltage versus input voltage for various values of bridge resistance. The results of these measurements are plotted in Figures 5, 6, and 7. For any desired bridge output voltage, these curves permitted the rapid choosing of the required lamps and bridge resistance as well as the required input voltage. From the curves it is apparent that there are two possible operating points for each lamp-resistor combination. One possible operating point is at point A, Figure 6, the center of the flat portion of the output voltage curve. For several volts change in the bridge voltage on either side of point A only a small change is apparent in E_o for some of the curves.

A voltage reference source was constructed with a lamp bridge operating in the center of the flat portion of the bridge output voltage curve. Two 6 watt-120 volt lamps were used because of the relatively flat output voltage characteristic obtained for a bridge resistance of 2000 ohms. The circuit constants are given in the schematic diagram of Figure 8. It soon became apparent, however, that, although the long term stability of the bridge was fairly good, the bridge was quite vulnerable to sudden changes in line voltage, since the bridge recovery time was strictly that of the lamps. An incremental response time of about 0.2 seconds for a slender 5 milliampere tungsten filament seems to be a fairly typical value.

For a complete stabilizer, Patchett[4] claims a response time of only one-tenth the incremental response time of the bridge elements. It is apparent that the stabilizer response time is a function of loop gain, and, for any reasonable amount of loop gain, will be considerably lower than the lamp response time. For this reason it was decided to try the second operating point of the lamp bridge.

2. Null-seeking lamp bridge

The second possible operating point for the lamp bridge is at

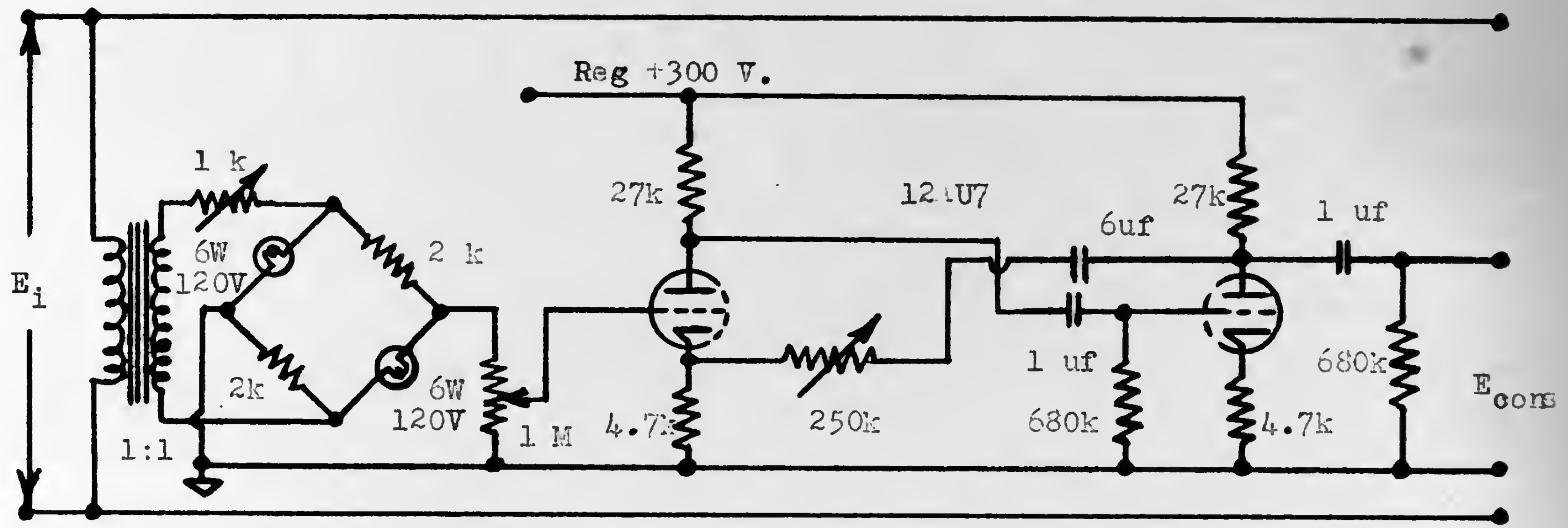


Figure 8. Lamp Bridge as A Constant Reference Voltage Source

point B, Figure 7, the bridge input voltage for which the output voltage is zero. There is actually a reversal in phase in E_o as the bridge voltage passes through point B; therefore all values of E_o above point B are plotted below the horizontal axis. Operation of the bridge at point B permits the derivation of an error signal that can be amplified and so connected in series with the input voltage that the output voltage seeks to return the bridge to its null output condition. It is under these conditions of operation that the unwanted bridge output voltages discussed previously become of great importance. Two 10 watt-250 volt lamps were used because of their greater immunity to vibration and shock than the other two types.

In the circuit of Figure 9, the magnitude of the bridge output error voltage is determined by the setting of the 1000 ohm bridge balance potentiometer. The error voltage is fed through a parallel T filter ($f_r = 180$ cycles), a phase shifter, two amplifier stages, a phase splitter, and then to the grids of the 807s. The secondary of the 807 output transformer is so connected as to oppose any change in E_i and thus maintain E_o constant.

In order to make the stabilizer function without regeneration or spurious oscillation, the parallel T filter was replaced by a low-pass filter with $f_o = 100$ cycles per second and $f_c = 180$ cycles per second. The filter stopped all spurious oscillations completely. Figure 10 shows the filter components and the attenuation and phase characteristics. For resistive loads (two 100 watt soldering irons) the stabilizer output waveform was good, and the output voltage varied only from 114.5 volts to 115.5 volts when the input voltage was varied between 100 and 132 volts. As soon as an instrument containing a power transformer was connected to the stabilizer, however, the third harmonic component in the transformer exciting current caused a voltage drop across the amplifier output transformer

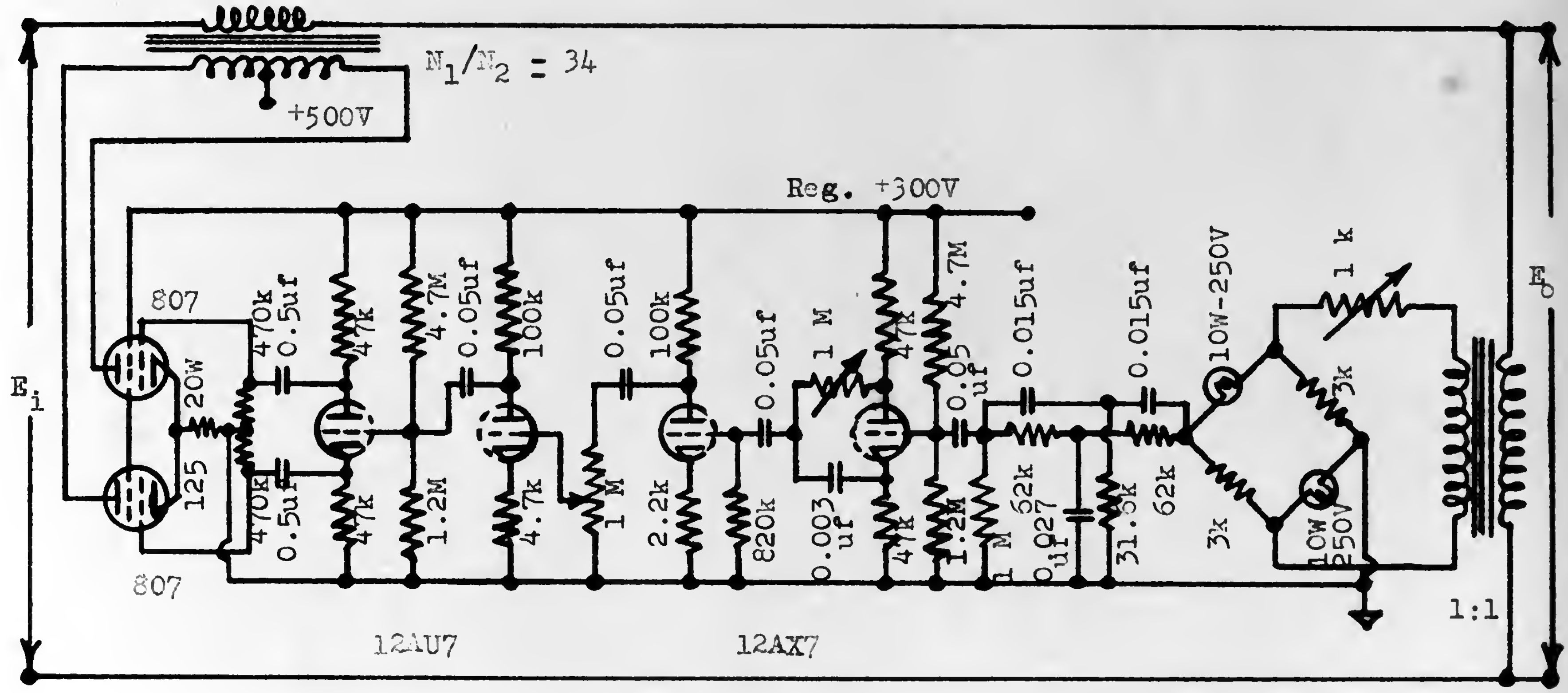


Figure 9. Null-Seeking Lamp Bridge and Third Harmonic Filter

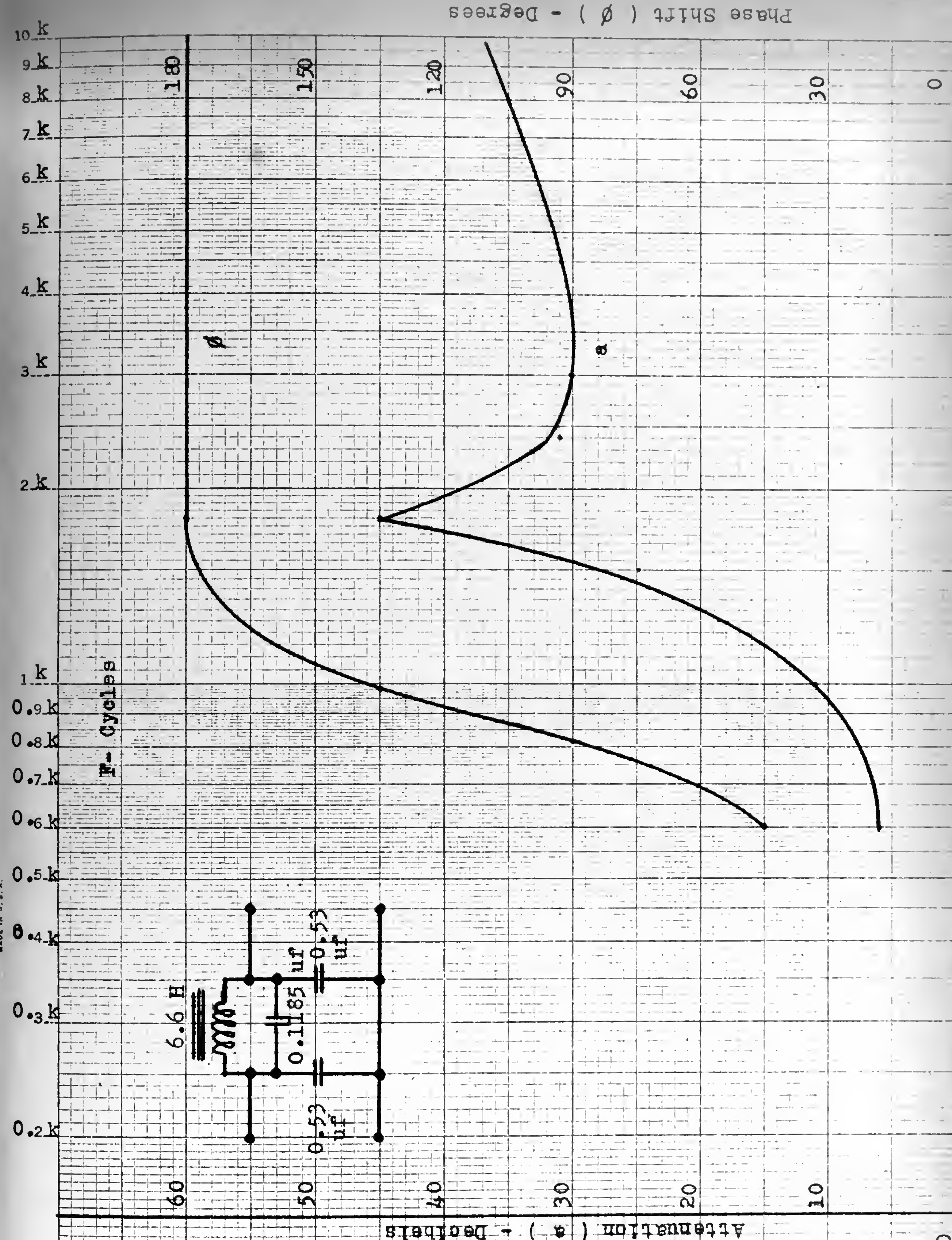


Figure 10. Low Pass Filter Characteristics

secondary, and this third harmonic voltage drop was subtracted from the stabilizer input voltage to give an output voltage waveform that was quite peaked and had pronounced shoulders. Since the low-pass filter between the bridge and the amplifier prevented the passage of third and higher harmonic components in the error voltage fed from the bridge detector to the amplifier, naturally no correction voltage could be developed across the amplifier output transformer secondary. Various forms of voltage feedback circuits were tried, in an attempt to reduce the load voltage waveform distortion, but any feedback circuit that reduced the waveform distortion at all also removed virtually all stabilizing action. For this reason, it was decided to try the third system of voltage stabilization, the details of which are discussed in Chapter III.

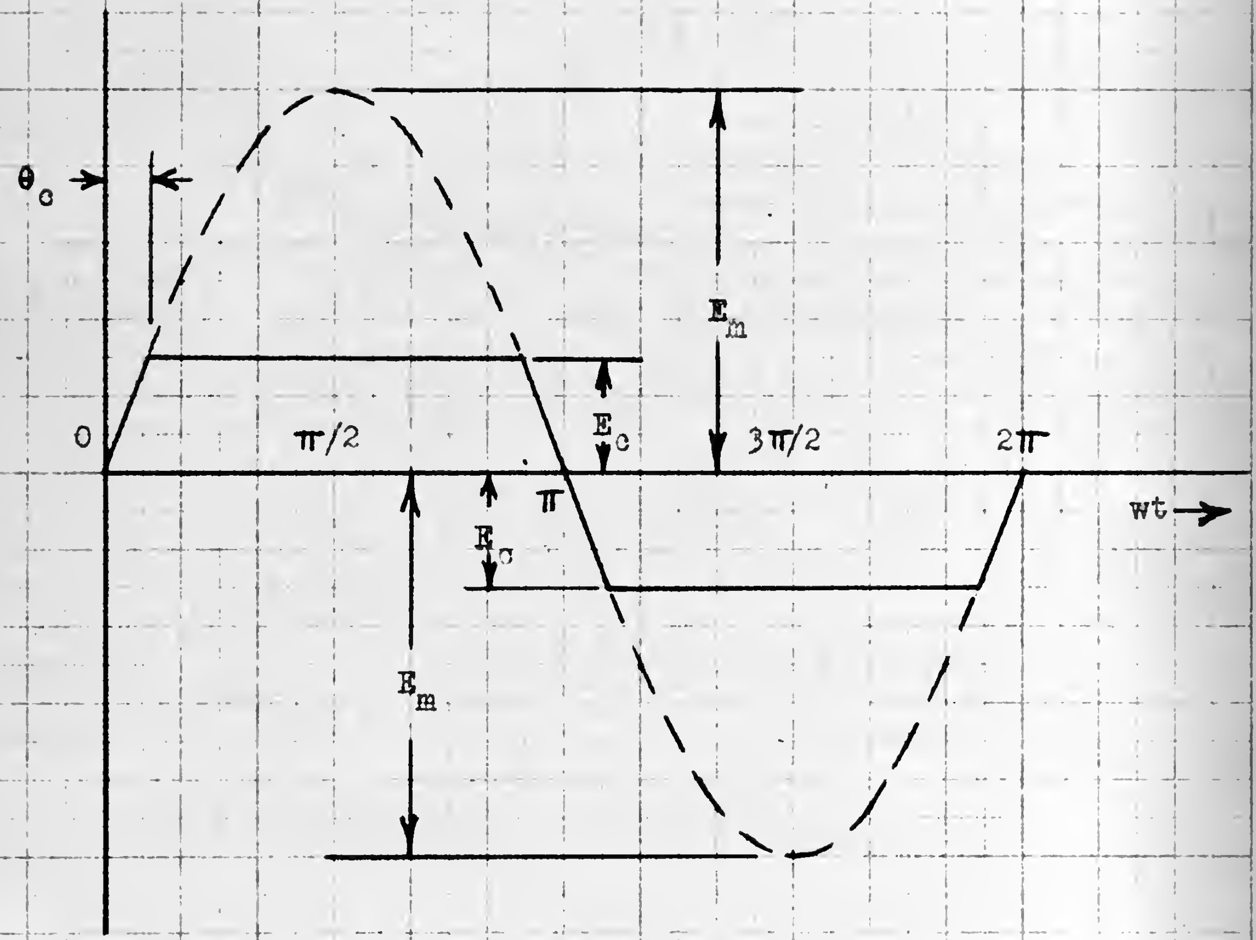


Figure 11. Constant Reference Voltage By Sine Wave Clipping

III. CONSTANT-LEVEL CLIPPING TECHNIQUES

The fundamental component of a square wave with an amplitude of E volts and frequency of $\omega/2\pi$ cycles per second is given by Terman[5] as: $E_f = \frac{4E}{\pi} \sin \omega t$. If, however, the sides of the square wave are not completely vertical, one would expect that the value of the fundamental component of the wave would be different from the value obtained for the square wave. For a heavily clipped sine wave, one would expect the value of the fundamental component to differ from that of the square wave, but to approach the value of the square wave E_f as the clipping angle approached zero degrees.

To show that the heavily clipped sine wave does offer a means of obtaining a constant reference voltage, a Fourier analysis* was made of the waveform shown in Figure 11 to determine the relation between E_f and θ_c, the clipping angle defined as: $\theta_c = \sin^{-1} E_c/E_m$. The results of this analysis give:

$$E_f = \frac{2\,E_c}{\pi} \left\{ \frac{\theta_c}{\sin \theta_c} + \cos \theta_c \right\}$$

It may be shown that $E_f \rightarrow \frac{4\,E_c}{\pi}$ as $\theta_c \rightarrow$ zero degrees and that $E_f \rightarrow E_c$ as $\theta_c \rightarrow$ ninety degrees, results that might have been expected from the previous discussion. To determine the necessary clipping angle for a desired degree of amplitude stability, values were computed for the expression:

$$\frac{\pi E_f}{2\,E_c} = \frac{\theta_c}{\sin \theta_c} + \cos \theta_c$$

*Appendix III

θ_c Degrees	θ_c Radians	$\frac{\theta_c}{\sin \theta_c} + \cos \theta_c$	% Deviation From Square Wave Value
0	0.00000	2.0000	0.00
1	0.01745	1.9998	0.01
2	0.03491	1.9997	0.015
5	0.08727	1.9975	0.13
10	0.17450	1.9900	0.50
20	0.34910	1.9606	1.98
30	0.52360	1.9132	4.34
40	0.69810	1.8521	7.40
50	0.87270	1.7820	10.90
60	1.04720	1.7092	14.50
70	1.22170	1.6420	17.90
80	1.39630	1.5916	20.40
90	1.57080	1.5708	21.50

Figure 12. Exact Values For $\frac{\theta_c}{\sin \theta_c} + \cos \theta_c$

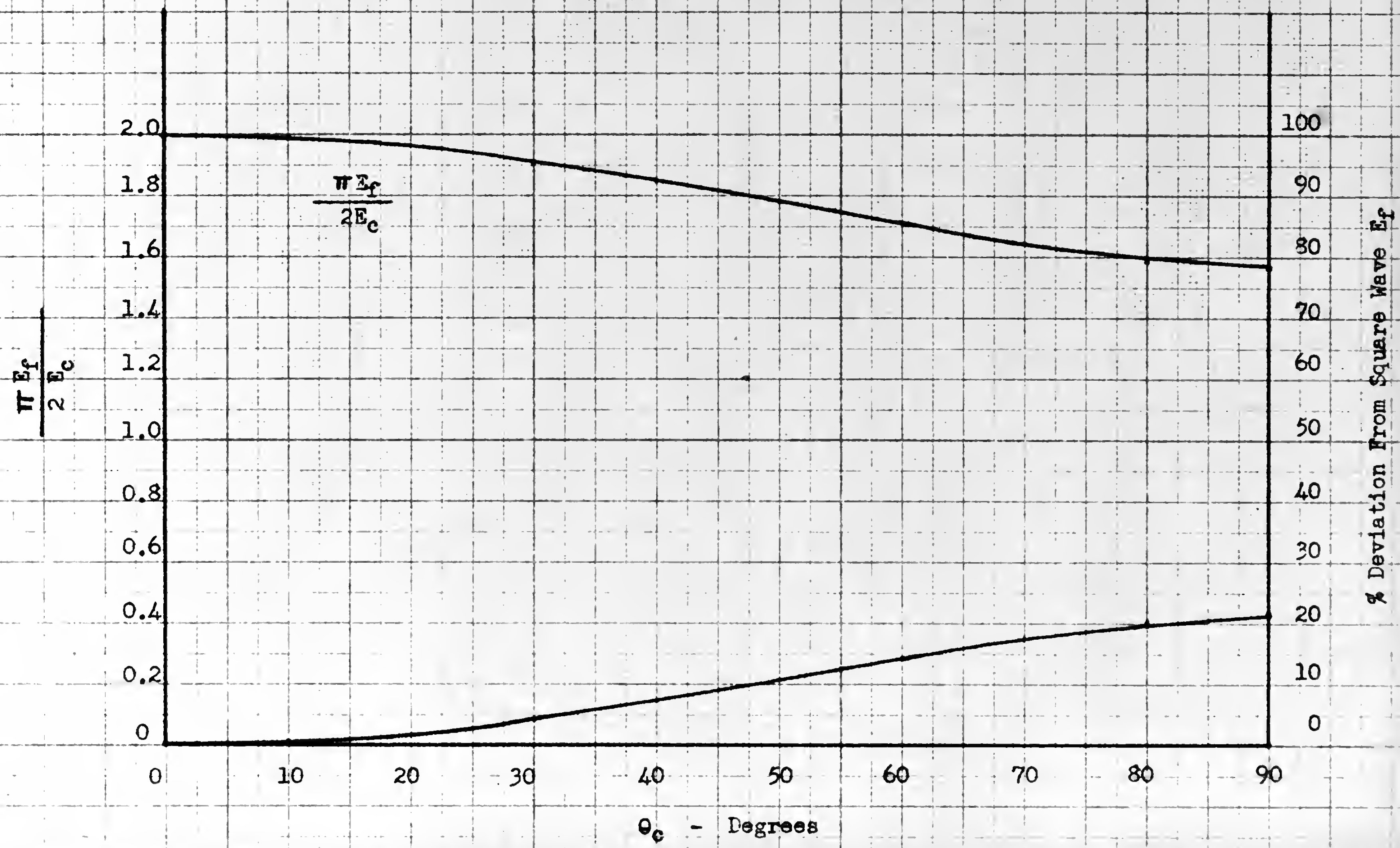


Figure 13. Variation of $\frac{\pi E_f}{2 E_c}$ With θ_c

when θ_c varied from zero degrees through ninety degrees. The exact values are tabulated in Figure 12 and plotted versus θ_c in Figure 13. The per cent deviation of E_f, for a given θ_c, from the square wave value of E_f is also tabulated and plotted in the same figures respectively.

From the tabulated values it is readily apparent that to obtain voltage stabilities of the order of 0.05%, the effective clipping angle for the allowable input voltage variation must be within the range from zero to three degrees.

In order to try out the possibilities of this system, a breadboard model of the voltage reference circuit was constructed, the final version of which is shown in Figure 14. This circuit involves stepping up the line voltage by a factor of seven, clipping the positive and negative excursions of the sine wave at fixed DC levels, and feeding the resulting wave through a low-pass filter to a phase-shifter and a voltage amplifier. In order to maintain the filter output constant over a fairly wide range in input frequency, care was taken to overcouple the two sections of the filter by means of the 1.47 uf mutual coupling capacitor and to load the output terminals of the filter quite heavily with a five thousand ohm potentiometer. Although the filter was originally designed as a symmetrical low-pass filter, it was found necessary to change the input capacitor from 0.5 uf to 0.22 uf to obtain the desired flat frequency response from fifty to seventy cycles per second.

The final version of the referency voltage source furnishes a zero to sixty-five volt (RMS) output that has no detectable variation on an HP 400C vacuum tube voltmeter as the input voltage is varied from ninety-five to one hundred and thirty-five volts at a given frequency. In addition, the output voltage varies only plus and minus 0.5 volts about a center

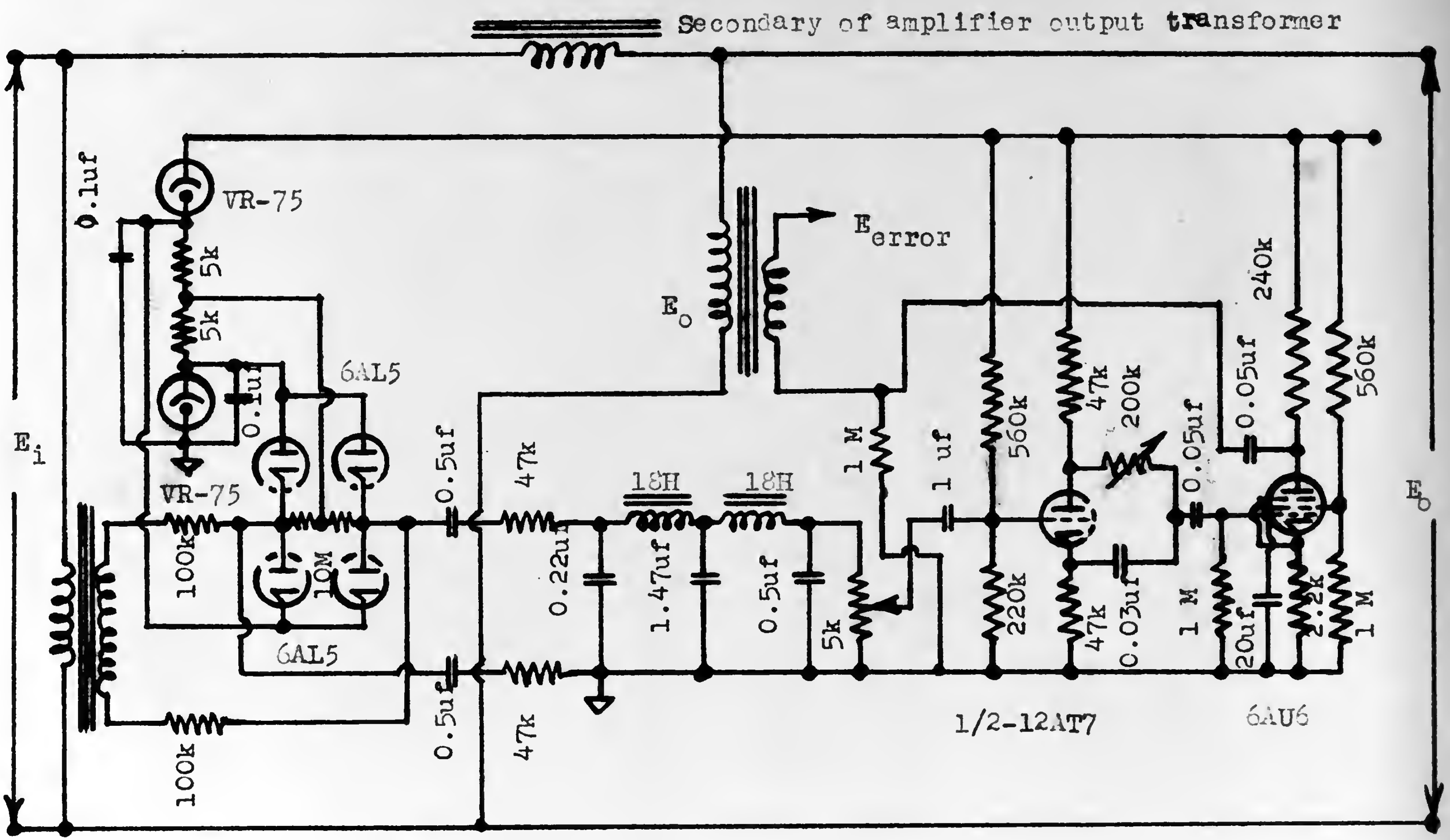


Figure 14. Final Schematic Of Clipped Sine Wave Reference Voltage Generator

frequency of sixty cycles as the input frequency is varied between fifty and seventy-four cycles per second. For an input frequency variation between fifty-five and sixty-five cycles per second, no variation in the voltage output can be detected. With the phase-shifter potentiometer in the worst position (0 ohms), the sixty volt output waveform contains 0.4% second harmonic distortion and 0.7% third harmonic distortion. Since the band-pass filter output waveform contains negligible second harmonic components and 0.5% third harmonic distortion, it is apparent that only a slight amount of second and third harmonic distortion is added by the output amplifier of the reference voltage generator.

IV. AMPLIFIER ANALYSIS

1. Stability criteria

Stability is one of the biggest problems in any stabilizer system employing an amplifier with negative feedback. For stabilizer stability, the vector gain around the loop, when plotted as a function of frequency, must not encircle the point (-1,0), corresponding to unity gain and 180 degrees phase shift. To allow a safety factor, it is normal practice to design the various circuits so that the phase shift around the entire loop will not be more than 150 degrees (corresponding to an amplitude-frequency characteristic slope of 10 decibels per octave) at any frequency for which the gain amplitude is greater than unity, and the gain amplitude will not be more than one-third for any frequency at which the phase shift is 180 degrees or more[6].

Figure 15 is a simplified version of the actual elements involved in the stabilizer feedback loop. In the stabilizer itself there are three elements that can cause phase shift: the amplifier, the amplifier output transformer, and the output voltage sampling transformer. In addition, the load itself, if highly reactive, can cause up to 90 degrees phase shift in the amplifier output circuit. For this reason it is essential that the two transformers be very high-quality units with low leakage reactance and distributed capacitance. Once the transformers have been selected, then the frequency response characteristic of the amplifier is the remaining parameter of the stabilizer.

Figure 16 is the schematic diagram for the error voltage amplifier in its final form. The design of the first two stages employs some of the techniques described by Terman[9] for feedback amplifiers. In brief, these techniques involve keeping the gain essentially constant over the

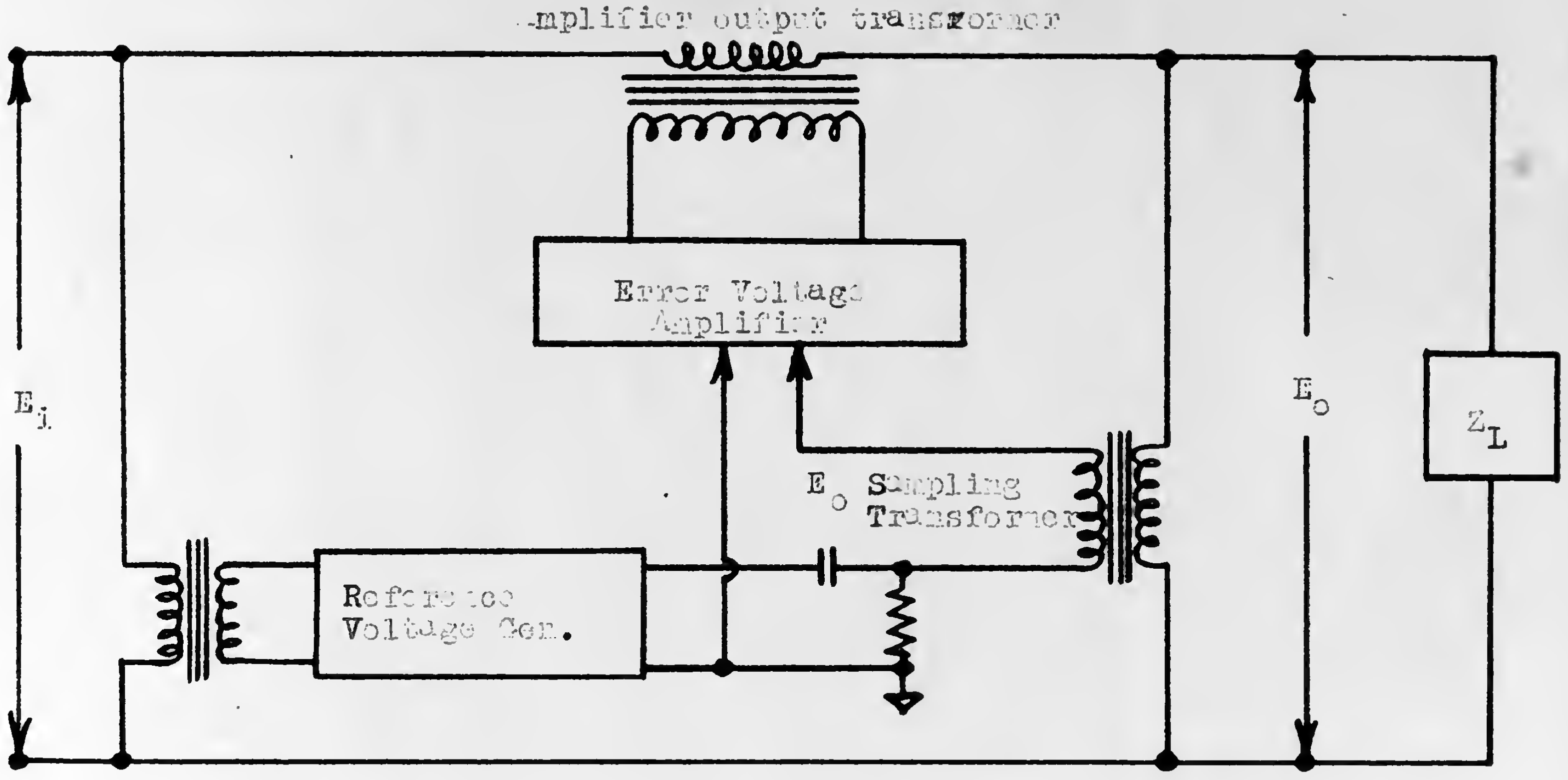


Figure 15. Stabilizer Feedback Loop Elements

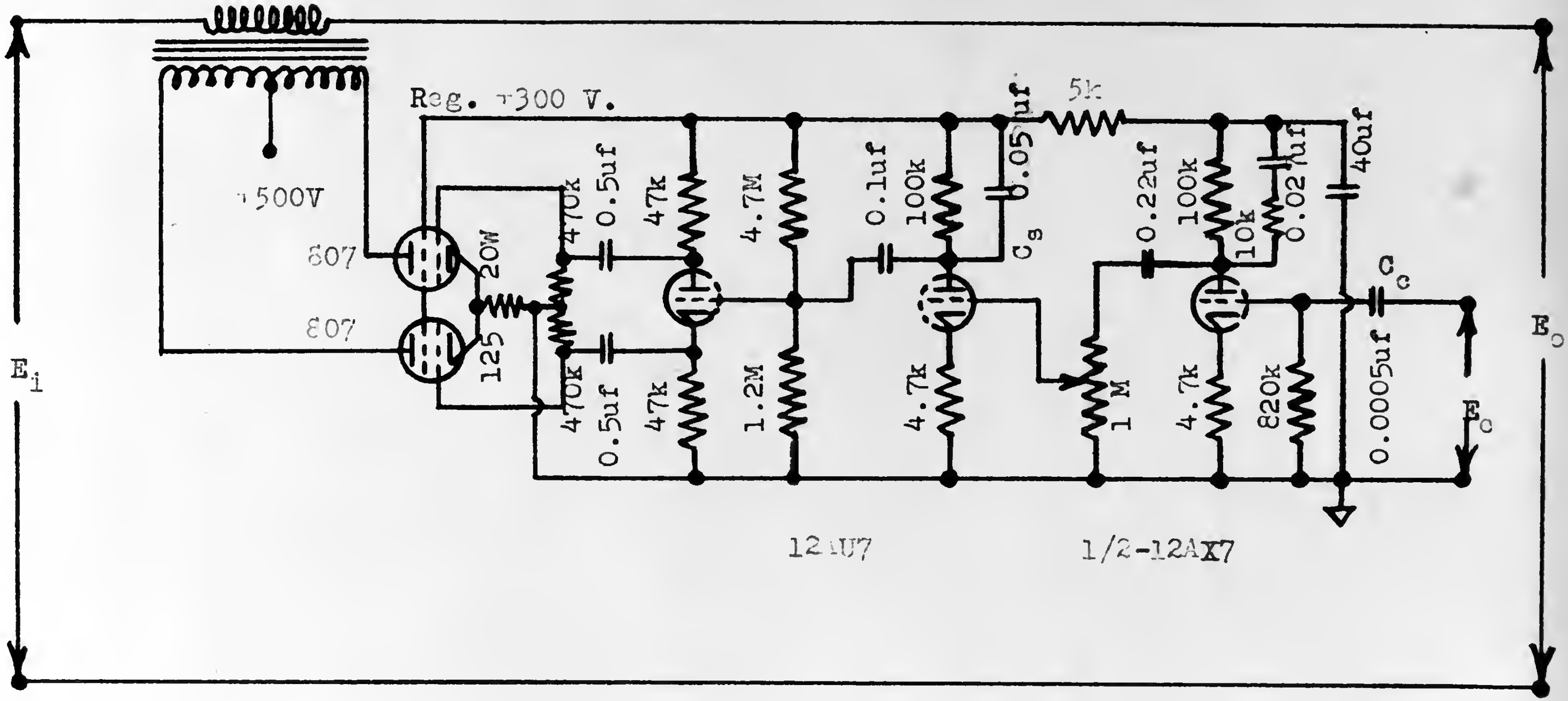


Figure 16. Final Schematic Of The Error Voltage Amplifier

desired frequency range and reducing the amplifier response by slightly less than 12 decibels per octave outside the desired frequency range. Terman[9] also shows that a gain-frequency slope of 12 decibels per octave corresponds to a phase shift of 180 degrees.

Figure 17 is the measured gain-frequency response for the amplifier of Figure 16. This curve was obtained by measuring the signal generator input voltage and the amplifier output voltage across a ten ohm, fifty watt resistor, rather than actually in the stabilizer loop. The fall-off in response at the low frequency end is primarily due to the 0.0005 uf input coupling capacitor which was necessary to prevent system motorboating. The fall-off in high frequency response is due to the plate load resistor bypass elements in the first and second stages. These elements were found necessary to prevent the system from oscillating at spurious higher frequencies.

2. Power amplifier analysis

The analysis of a push-pull amplifier delivering power to a load has been covered in considerable detail in the literature. The operating conditions for a push-pull amplifier absorbing power are not nearly so well-known. Clapp[6] and Patchett[7] have both analyzed the push-pull amplifier for the power-absorbing condition, and the end results obtained by the two methods agree reasonably well. In order to realize the maximum power output from the amplifier, it would seem desirable to operate the output tubes class B. Clapp[6], however, claims that better performance, with respect to sinusoidal output waveform under adverse load and line voltage waveforms, is obtained when the output stage is operated class AB_1 with fixed bias. Patchett[7] states that the bypassed cathode resistor of the output stage

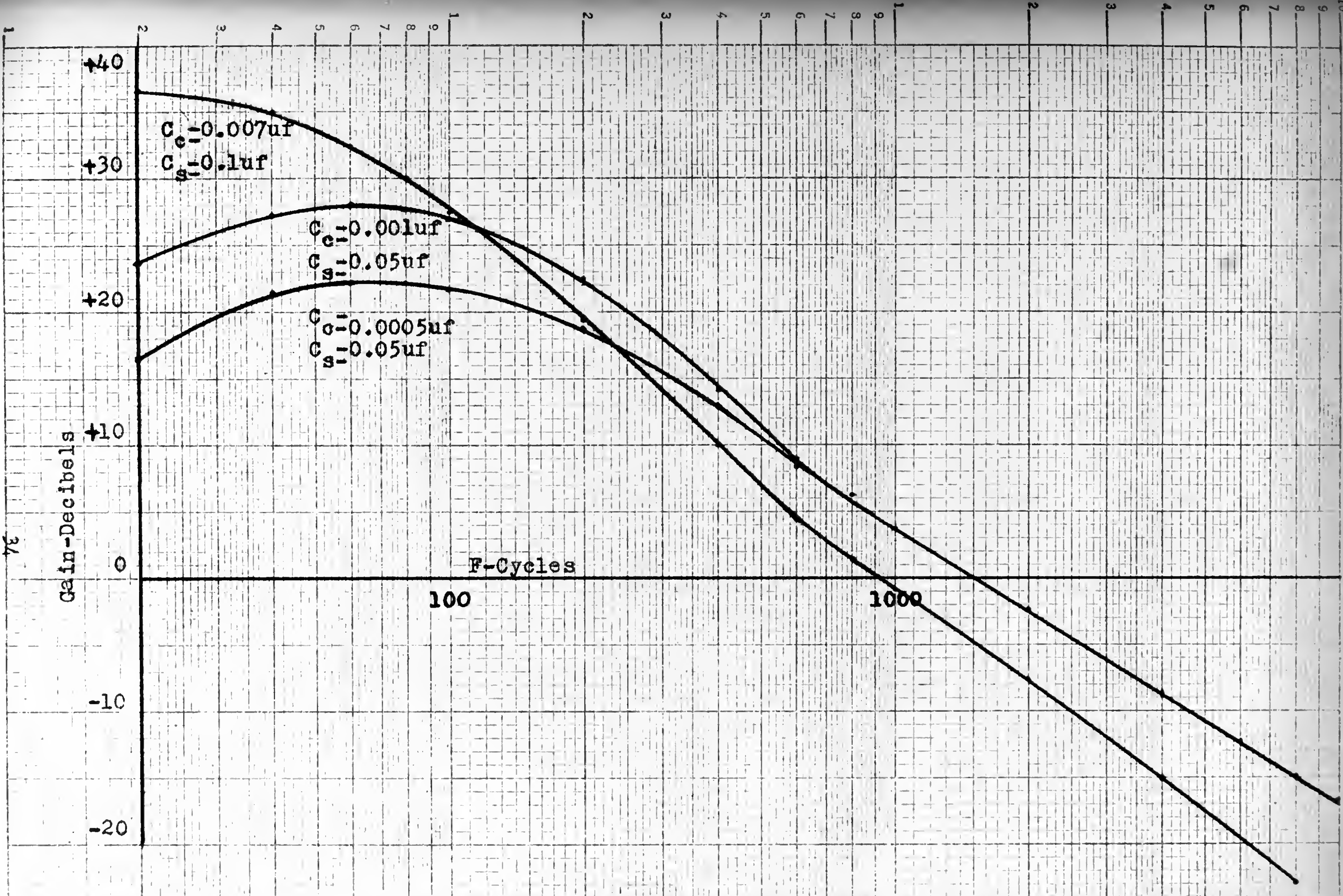


Figure 17. Measured Gain VS Frequency Characteristics Of The Error Voltage Amplifier

is a convenient means of reducing the regulation required of the plate voltage power supply, since the output stage bias changes as the stabilizer operating conditions change. Mak[8] claims that the distortion in the amplifier output has very little effect upon the stabilizer output waveform because of the relative magnitudes of the quantities involved and the feedback action of the stabilizer.

A brief resume of the output amplifier analysis given by Clapp[6] is included below. For greater detail, the original reference may be consulted.

Figure 18 shows the E_b vs I_b curve for an 807. Only the $E_c = 0$ curve is shown to avoid confusion. For a conventional push-pull amplifier delivering power to a load, circuit conditions may be analyzed graphically by means of a load line drawn on the characteristic curves for the output tubes. Under these conditions the load line relates, for any control grid voltage, the plate current and plate voltage as functions of the initial operating conditions and the impedance developed in the plate circuit of the tubes by the output load. For this situation, the important variable of operation is the control grid voltage.

For the stabilizer, operating as shown in Figure 15, the output voltage is fixed, while the load current and the input voltage are variables. It is necessary that the grids of the output tubes be driven so that, when the specified I_L is flowing, the voltage across the output transformer secondary satisfies the relation: $E_a = E_o - E_i$, where due regard is taken of the algebraic sign of E_a. The volt-ampere output of the stabilizer is therefore:

$$W_a = I_L E_a = I_L E_o - I_L E_i$$

As the stabilizer input voltage changes from its minimum value to its maximum value, the stabilizer volt-ampere output changes by an amount:

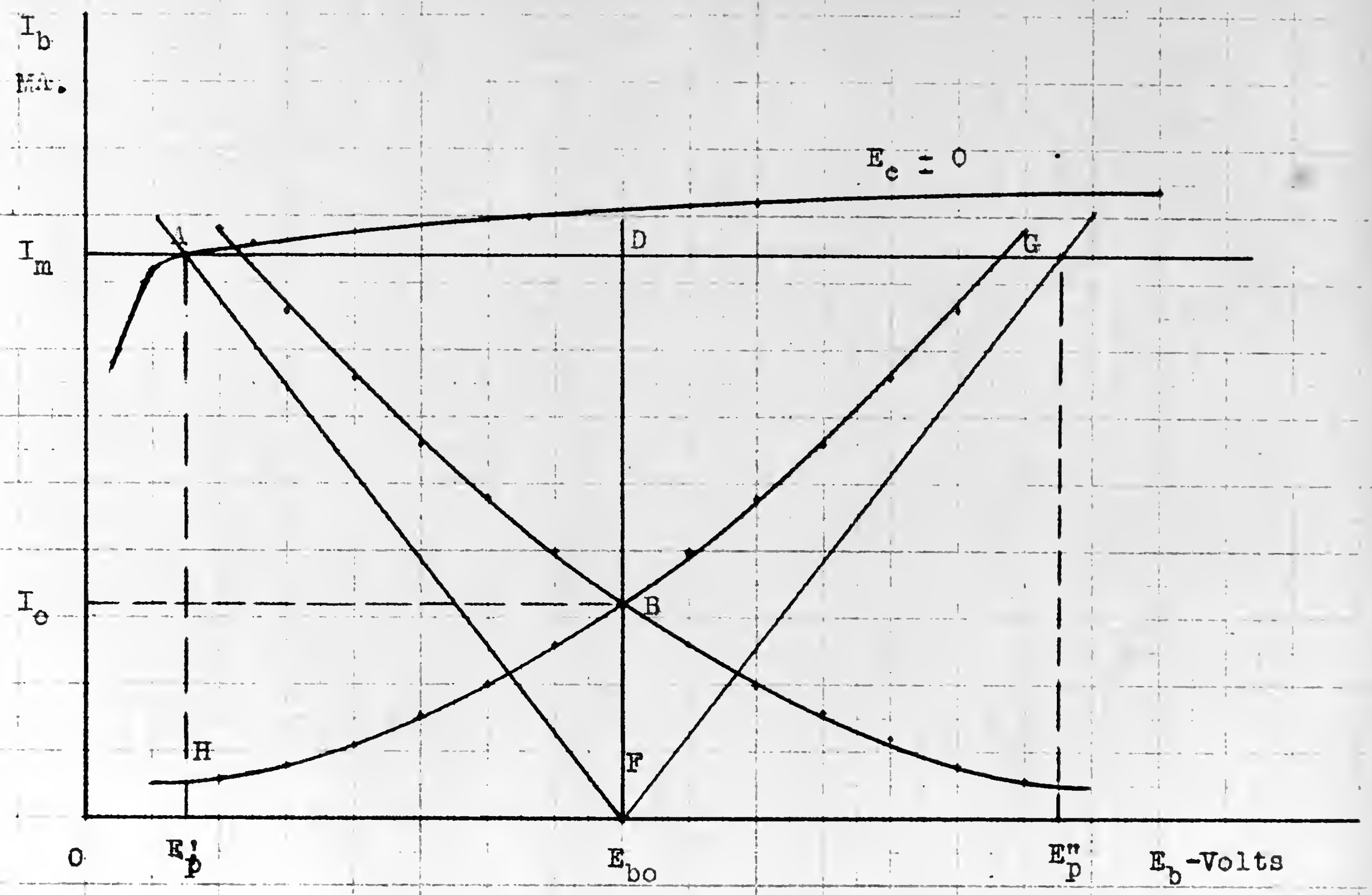


Figure 13. Characteristic Curve For An 807 With $E_c = 0$

$$W_a' - W_a'' = (E_i'' - E_i')\, I_L$$

where: E_i' = minimum input voltage

E_i'' = maximum input voltage

The change in output volt-amperes is a measure of the effectiveness of a stabilizer, similar to the power output rating of a conventional amplifier. For output tubes operating class B, Patchett(7) obtains the relation:

$$W_a' - W_a'' = 0.77\, W_d$$

where: W_d = maximum allowable plate dissipation for the output tubes

Therefore, for two 807's operated class B, assuming $\pm$ 10% variation in E_i with E_o constant, the maximum load = 188 volt amperes. Load line ABC in Figure 18 is a conventional load line for one tube of a push-pull amplifier, operating class A, that is delivering power to a resistive load. It represents the condition of maximum load current I_L and minimum input voltage E_i. When $E_i = E_o$, $E_a = 0$, and for this condition the load line is DBF. When the input voltage reaches its maximum value, the load line shifts to EBG, representing power delivered from the power line to the plates of the amplifier output tubes. This is the limiting condition of operation for the stabilizer, since for it the output tube plate dissipation is a maximum.

For one of the push-pull tubes whose characteristics are shown in Figure 18, considering only the 60 cycle components of current and voltage, we may write:

$$W_p = E_{bo} I_o - \frac{1}{2} (E_{bo} - E_m)\, I_m$$

where: I_o = average or DC component of the tube plate current

E_m = value of the AC plate voltage at the time when I_p has its maximum value

$$\left(E_m = E_{bo} - E_p' = E_p'' - E_{bo} = \frac{E_a}{2} \frac{N_1}{N_2} \right)$$

I_m = maximum or peak value of the fundamental component of the plate current

$$\left(I_m = \frac{2 N_2 I_r}{N_1}\right)$$

Rearranging this expression gives:

$$W_p = E_{bo}\left(I_o - \frac{I_m}{2}\right) - \frac{E_m I_m}{2}$$

To select the best operating conditions for the stabilizer output tubes, one would like to obtain the maximum possible volt-ampere rating while keeping W_p below the maximum safe value of plate dissipation permitted for the output tube type. Obviously, from the relation above, W_p is a minimum when $E_{bo} = 0$, however, Clapp[6] found that using the output tubes as a variable power absorber prevents the unit from functioning as a waveform stabilizer. For this reason, Clapp operated his output tubes class AB_1, determining the operating conditions in the following manner:

1. Choose point A at the knee of the zero bias curve for the output tube. This determines E'_p and I_m.
2. By trial, select point C (and E''_p) such that, with $E_{bo} = \frac{1}{2}(E'_p + E''_p)$, W_p is less than the safe allowable plate dissipation for the class of service.
3. With the amplifier delivering its maximum output power (operating along load line ABC), check to make sure that the allowable screen dissipation is not exceeded.

V. CONCLUSIONS AND RECOMMENDATIONS

The reference voltage generator of Figure 14 was connected to a 60 cycle power line bus that was fed from the commercial power outlet through a variac. By varying the amplitude control potentiometer and the phase shift potentiometer the fundamental component of the sample line voltage could be nulled out, leaving only the higher harmonic voltages as an error signal output from the reference voltage unit. If a higher stabilizer output voltage is desired, the amplitude control of the reference voltage generator is turned to raise the output voltage of the unit and a fundamental component will also then be present in the error voltage output. The reverse procedure applies if a lower stabilizer output voltage is desired.

The error signal was then fed into the error voltage amplifier, the output of which was placed in series with the stabilizer input voltage. Two types of load were used with the stabilizer: a 110 watt soldering iron and an HP 410B vacuum tube voltmeter. The power consumption for the HP410B is given in the maintenance manual as approximately 40 watts. Although it does not utilize the full control rating of the stabilizer, the instrument does contain a power transformer, and thus gives a test of the stabilizer performance with devices containing power transformers. The results obtained for the two loads are tabulated in Figure 19.

It will be noted that, for frequencies inside the pass band of the error voltage amplifier, there is a marked improvement in the stability and the harmonic content of the load voltage. The average Stabilization Ratio is equal to, or better than the full load value of 50, where $S \equiv \frac{(\delta V_i)(V_o)}{(\delta V_o)(V_i)}$. A reduction in the input voltage harmonic content by

	AC LINE	HP410B LOAD			110 WATT RESISTIVE LOAD		
E_{in}	115 v.	105 v.	115 v.	125 v.	105 v.	115 v.	125 v.
E_{out}	115 v.	114.9	115	115.1	114.8	115	115.2
H_2	---	0.37%	0.47%	0.51%	0.43%	0.48%	0.52%
H_3	3.7%	0.36%	0.41%	0.69%	0.59%	0.11%	0.51%
H_4	---	--	--	--	--	--	--
H_5	0.60%	0.60%	0.15%	--	0.79%	--	0.13%
H_6	---	--	--	--	--	--	--
H_7	0.30%	0.58%	0.12%	0.19%	0.64%	--	0.11%
H_8	---	--	--	--	--	--	--
H_9	0.25%	0.21%	0.17%	0.15%	0.31%	--	--
H_{10}	---	--	--	--	--	--	--
H_{11}	0.15%	--	--	--	--	--	--
H_{12}	---	--	--	--	--	--	--
H_{13}	0.14%	--	--	--	0.14%	--	--
H_{14}	---	--	--	--	--	--	--
H_{15}	0.20%	--	--	--	0.13%	--	--

(Readings were taken only for those harmonics $>$ 0.1%)

Figure 19. Stabilizer Performance With Two Different Types of Loads

an average factor of 10 may also be noted in the columns of Figure 19. The effective internal resistance of the stabilizer is approximately:

$$R_o = \frac{-1.0}{1.0} = -1.0 \text{ ohm.}$$

In order to check the stabilizer recovery time for a sudden change in the input voltage, a 10 ohm 50 watt resistor was placed in series with one side of the AC line from the variac to the stabilizer. By means of a toggle switch the resistor could be removed from or inserted in the line as desired. Simultaneous Brush recordings were made of the input and output voltages of the stabilizer while a 110 watt soldering iron was connected across the stabilizer output. The stabilizer input voltage was set at various values by means of the variac, and then sudden changes in line voltage were simulated by operation of the resistor shorting switch. Figure 20 is a reproduction of the input and output voltage waveforms. The step in the lower waveform represents an eleven volt change in the input voltage. No detectable change is present in the output voltage waveform, so that the stabilizer action is essentially instantaneous. Further expansion of the amplitude of both waveforms was found impossible because of the recording characteristics of the Brush recorder used.

1. Conclusions

The theoretical aspects of the stabilizer are substantiated by the actual model that was built and tested. The final form of the stabilizer, using a clipping technique to obtain a constant reference voltage, seems to offer greater possibilities for a precision stabilizer than either of the other two systems tried.

Careful control of the vector gain or the phase shift around the stabilizer loop is necessary to insure system stability. This control

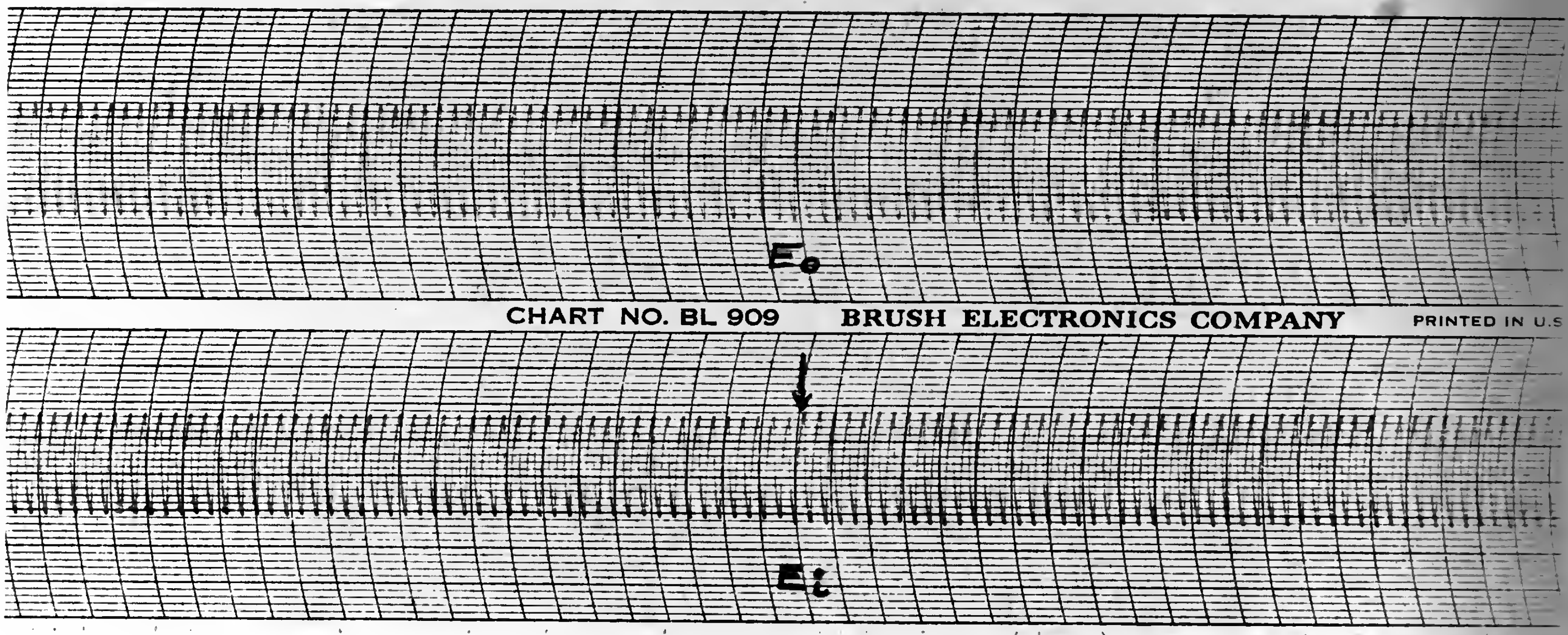


Figure 20. Stabilizer Recovery Time For Sudden Input Voltage Changes

should include the selection of high-quality transformers and the shaping of the amplifier response characteristic to meet the requirements of the Nyquist criteria for stability.

2. Recommendations

The overall performance of the stabilizer can be no better than the reference voltage generator with which the output voltage and waveform are compared. From the analysis of the clipped sine wave in Chapter III, it appears that an improvement in stability may be achieved by reducing the effective clipping angle. One method of reducing θ_c is by the use of an overdriven amplifier. Another method of decreasing the effective clipping angle is by use of the cathode-coupled clipper shown in Figure 21. For proper selection of the circuit constants, V1 and V2 will both be conducting when there is no AC input signal to the grid of V1 For an AC input to the grid of V1 as shown, the increase in V1 plate current through R_k is large enough to cut V2 off. When the AC input voltage goes negative enough to cut V1 off, then V2 conducts again. For R_L and R_k fairly large, the output voltage will be almost a square wave. If further squaring of the wave is desired to reduce the effective clipping angle, a second cathode-coupled clipper may be employed.

The stability of the output voltage is dependent upon the constancy of the cutoff values of V1 and V2, as well as the regulation of the regulated plate supply. Since the variation of the plate supply voltage can be made less than 0.01% without undue complexity of regulating circuitry, the stability of the proposed system then becomes almost solely dependent upon the cutoff values of V1 and V2. The tube cutoff potential is a function of the physical position of the tube elements as well as the plate

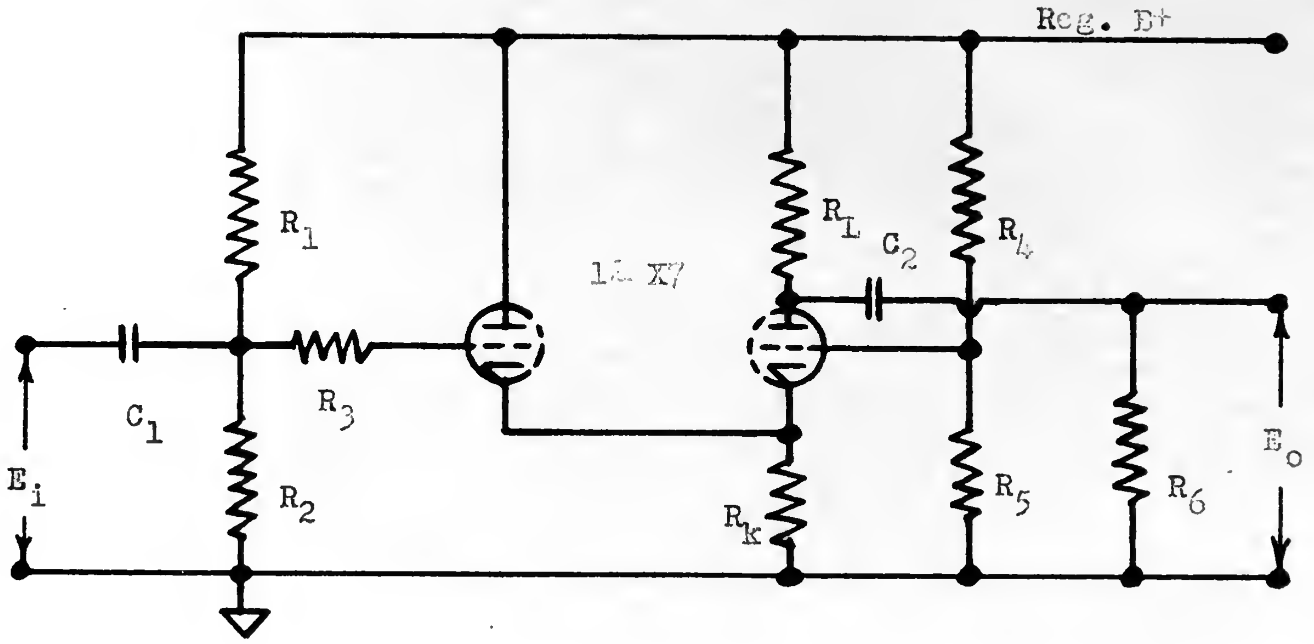


Figure 21. Cathode-Coupled Sine Wave Clipper

voltage. For a fixed plate voltage, the cutoff potential is relatively constant, if the tube is not subjected to excessive shock or vibration. For this reason, the cathode-coupled clipper is recommended as one means of improving the stability of the system.

One very vital requirement for a stabilizer is that it be capable of controlling sufficient power to make its cost, complexity, and space consumption worthwhile. The stabilizer that was constructed and tested is conservatively estimated to be capable of controlling 150 volt-amperes with input voltage variations from 105 to 125 volts and load power factor variations from 0.5 lead to 0.5 lag. It is possible to increase the volt-ampere capacity of the system by a factor of 20 to 30, increasing the cost by about a factor of 2, if the system proposed in Figure 22 is employed. This system utilizes a second winding on the error voltage amplifier output transformer to excite the control field of a two phase 60 cycle servo motor. The reference field of the motor is excited from the stabilizer input terminals through a 10 uf capacitor to produce the desired quadrature reference field in the motor. When the stabilizer input voltage changes, the error voltage amplifier produces an output voltage that bucks the change and attempts to keep E_o constant. At the same instant, the error voltage is fed to the control field of the servo motor. The shaft of the servo motor is connected through a gear train to the arm of a variac whose outside terminals are connected across the line as shown. Motion of the variac arm causes a net buck or boost voltage to appear across the secondary winding of the series transformer, thus bringing the stabilizer input voltage back to its original value.

For proper selection of components, the instantaneous line voltage variations could be compensated for by the amplifier output, while long

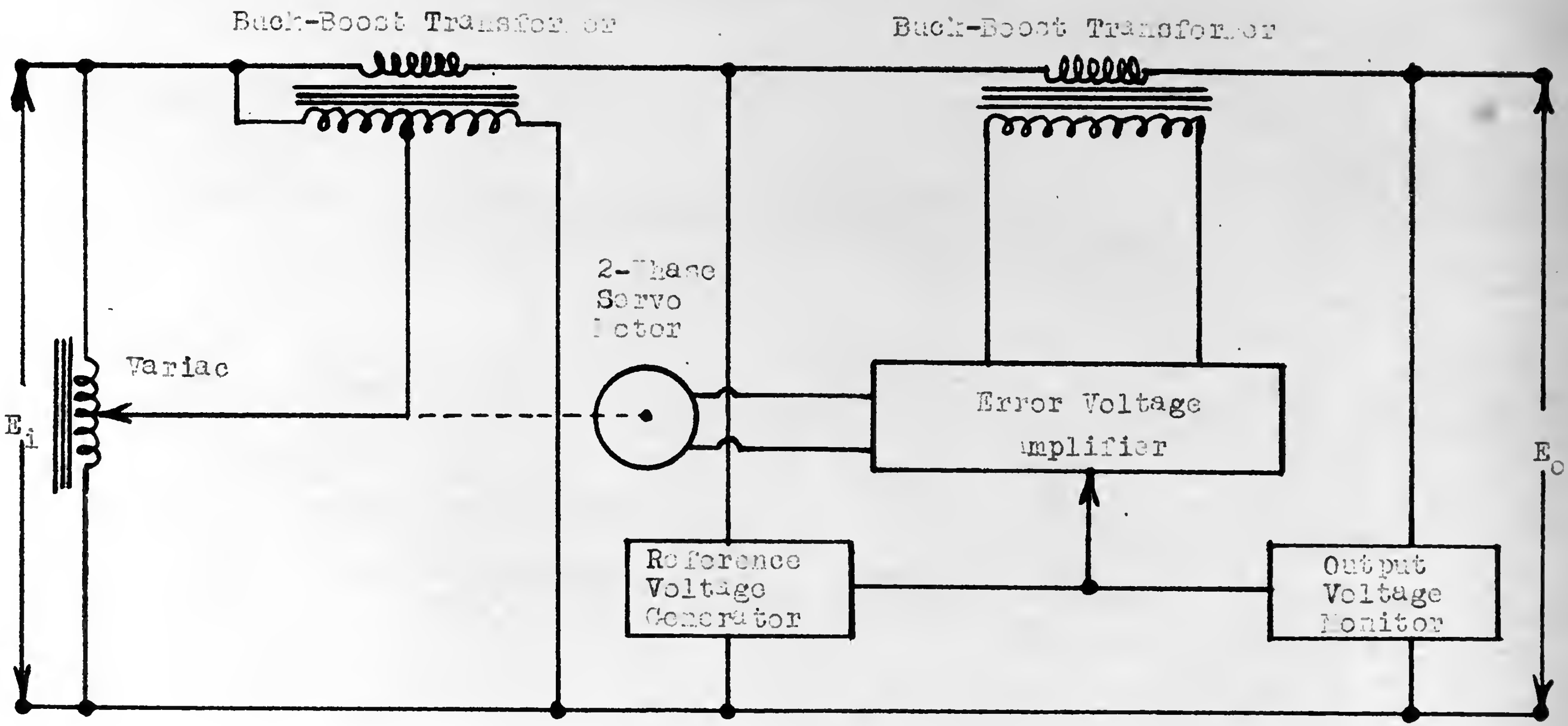


Figure 22. Proposed System For Increasing The Stabilizer Power-Handling Capacity

term variations are corrected by the fast-acting servo loop. This system is recommended as one possible means of greatly increasing the stabilizer volt-ampere rating.

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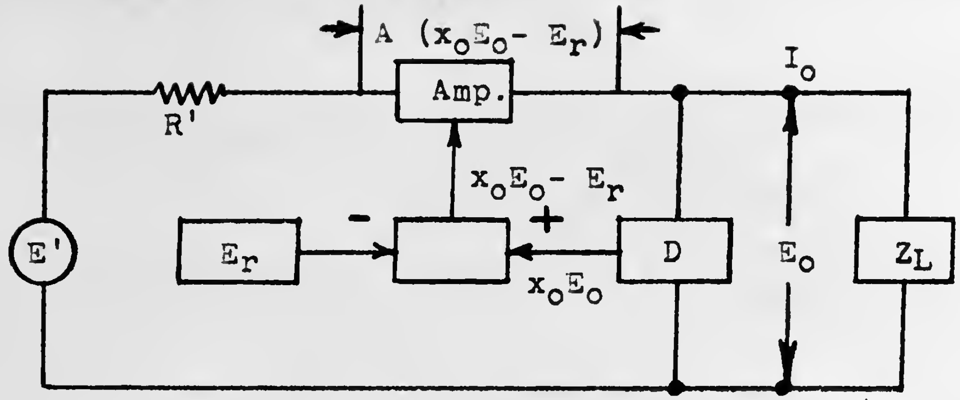
APPENDIX I



Using the circuit shown above, where the symbols are as defined on page 7, we may write:

$$E' = I_oR' + A_o\,(x_oE_o - E_r) + E_o$$

$$E' = I_oR' + A_ox_oE_o - A_oE_r + E_o$$

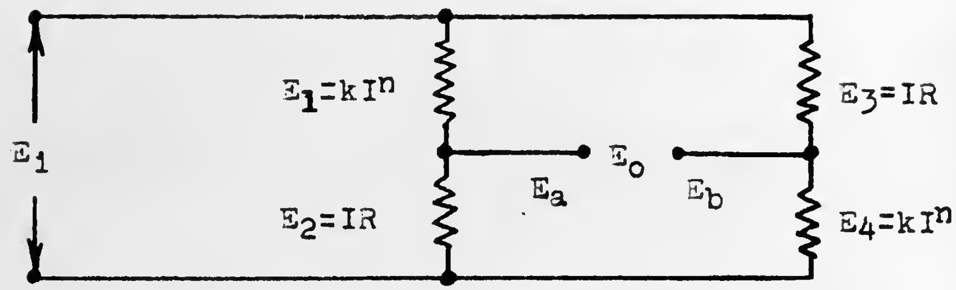
$$E' = I_oR' + (1 + A_ox_o)\,E_o - A_oE_r$$

Taking differentials on both sides:

$$0 = dI_oR' + (1 + A_ox_o)\,dE_o$$

$$-dI_oR' = (1 + A_ox_o)\,dE_o$$

$$R_o \equiv \frac{dE_o}{dI_o} = \frac{-R'}{(1 + A_ox_o)}$$



For the circuit above we may write:

$$E_a = E_i - kI^n$$

$$E_b = E_i - IR$$

$$E_o = E_b - E_a$$

$$E_o = E_i - IR - (E_i - kI^n)$$

$$E_o = kI^n - IR$$

Taking the derivative of both sides with respect to E_i :

$$\frac{dE_o}{dE_i} = knI^{n-1}\left(\frac{dI}{dE_i}\right) - R\left(\frac{dI}{dE_i}\right)$$

For either leg of the circuit,

$$E_i = IR + kI^n$$

$$1 = \frac{RdI}{dE_i} + knI^{n-1}\left(\frac{dI}{dE_i}\right)$$

$$\frac{dI}{dE_i} = \frac{1}{R + knI^{n-1}}$$

$$\therefore \quad \frac{dE_o}{dE_i} = \left[knI^{n-1} - R\right]\left[\frac{1}{R + knI^{n-1}}\right]$$

At the balance point of the bridge, $E_o = 0$

$$E_o = kI^n - IR = 0$$

$$IR = kI^n$$

$$R = kI^{n-1}$$

$$\left.\frac{dE_o}{dE_i}\right]_{E_o = 0} = \frac{Rn - R}{Rn + R}$$

$$\left.\frac{dE_o}{dE_i}\right]_{E_o = 0} = \frac{n - 1}{n + 1}$$

APPENDIX III

Using the nomenclature of Figure 11, and assuming that perfect clipping occurs at voltage level E_c, where $E_c \equiv E \sin \theta_c$:

$$E_f = \frac{2}{\pi} \left\{ \int_0^{\theta_c} (E \sin x) \sin x \, dx + \int_{\theta_c}^{\pi-\theta_c} (E \sin \theta_c) \sin x \, dx + \int_{\pi-\theta_c}^{\pi} (E \sin x) \sin x \, dx \right\}$$

$$E_f = \frac{2E}{\pi} \left\{ \int_0^{\theta_c} \sin^2 x \, dx + \int_{\theta_c}^{\pi-\theta_c} (\sin \theta_c) \sin x \, dx + \int_{\pi-\theta_c}^{\pi} \sin^2 x \, dx \right\}$$

$$E_f = \frac{2E}{\pi} \left\{ \left[\frac{x}{2} - \frac{\sin 2x}{4} \right]_0^{\theta_c} + \sin \theta_c \Big[-\cos x \Big]_{\theta_c}^{\pi-\theta_c} + \left[\frac{x}{2} - \frac{\sin 2x}{4} \right]_{\pi-\theta_c}^{\pi} \right\}$$

Substituting limits and combining:

$$E_f = \frac{2E}{\pi} \left\{ \theta_c + \frac{\sin 2\theta_c}{2} \right\}$$

Since $E = \dfrac{E_c}{\sin \theta_c}$

$$E_f = \frac{2E_c}{\pi} \left\{ \frac{\theta_c}{\sin \theta_c} + \cos \theta_c \right\}$$

For $\theta_c \rightarrow 0^\circ$:

$$E_f = \lim_{\theta_c \to 0} \left\{ \frac{2\,E_c}{\pi} \left[\frac{\theta_c}{\sin \theta_c} + \cos \theta_c \right] \right\}$$

$$E_f = \frac{2\,E_c}{\pi} \left\{ \frac{1}{\cos 0} + \cos 0 \right\}$$

$$E_f = \frac{2\,E_c\,(2)}{\pi} = \frac{4\,E_c}{\pi}$$

For $\theta_c \rightarrow \frac{\pi}{2}$:

$$E_f = \lim_{\theta_c \to \frac{\pi}{2}} \left\{ \frac{2\,E_c}{\pi} \left[\frac{\theta_c}{\sin \theta_c} + \cos \theta_c \right] \right\}$$

$$E_f = \frac{2\,E_c}{\pi} \left\{ \frac{\pi/2}{\sin \pi/2} + \cos \pi/2 \right\}$$

$$E_f = \frac{2\,E_c}{\pi} \left\{ \frac{\pi}{2} \right\} = E_c$$

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